10/590671
IAP5 Rec'd PCT/PTO 24 AUG 2006

CONSTRAINED OPTIMIZATION BASED MIMO LMMSE-SIC
RECEIVER FOR CDMA DOWNLINK

FIELD OF THE INVENTION

[0001] This invention pertains in general to communication systems. More particularly, embodiments of the invention pertain to transmit diversity and Multiple-In, Multiple-Out (MIMO) transmission and receiving methods for multiple antenna technology of Code Division Multiple Access (CDMA) type systems.

BACKGROUND

[0002] Inter-chip interference (ICI) is a result of the multipath frequency selective channel in the CDMA downlink. The presence of ICI destroys the orthogonality of the Walsh spreading codes at mobile terminals. The challenge to the receiver design is even greater for a MIMO system in the CDMA downlink. The receiver has to combat both the ICI and the co-channel interference (CCI) to achieve reliable communication. Therefore, interference cancellation at the mobile stations is an effective means of improving the receiver performance and link capacity.

[0003] For a Single-In, Single-Out (SISO) or Single-In, Multiple-Out (SIMO) system, there are two types of linear equalizers: non-adaptive and adaptive equalizers. A non-adaptive equalizer usually requires matrix inversion and is therefore computationally expensive (I. Ghauri and D. T. M. Slock, "Linear Receivers for the DS-CDMA Downlink Exploiting

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Orthogonality of Spreading Sequences" in Proc. of 32nd Asilomar

Conference, pages 650 – 654, 1998; A. Klein, "Data Detection Algorithms

Specially Designed for the Downlink of CDMA Mobile Radio Systems" In

Proc. of VTC 97, pages 203-207, Phoenix AZ, 1997; T. P. Krauss, W. J.

Hillery and M. D. Zaltowski, "MMSE Equalization for forward link in 3G

CDMA: Symbol Level Versus Chip-level" In Proc. of 10th IEEE Workshop

on Statistical Signal and Array Processing, pages 18 – 22, 2000; S.

Werner andJ. Lilleberg, "Downlink Channel Decorrelation in CDMA

Systems With Long Codes", In Proc. of 49th VTC, pages 1614 – 1617,

1999), especially when the coherence time of the channel is short and the

equalizer requires frequent update. An adaptive equalizer is less

computationally involved since it does not require direct matrix inversion,

but it is not as robust as the non-adaptive equalizers since its performance

is sensitive to the choice of parameters such as step size, initialization, etc

(M. J. Heikkila, "A Novel Blind Adaptive Algorithm for Channel

Equalization in WCDMA Downlink", In Proc. of 12th IEEE Symposium or

PIMRC, pages A41 – A45); (W. J.Heikkil/a, P, Kowu/aises, and J.

Lilleberg, "Interference Suppression in CDMA Downlink Through

Adaptive Channel Equalization" In Proc. of VTC 99-Fall, pages 978 – 982,

Amsterdam,TheNetherlands, 1999); (Mika Ventola, "Performance Analysis

of Dual ANtenna Reception in WCDMA Downlink: issue 1,0" NMP/RTA,

March 2002). Solutions to these difficulties have been proposed (J.

Zhang, T. Bhatt, and G. Mandyam, "Efficient FFT Based Linear

Equalization for CDMA Downlink", in Proceedings of Asilomar
Conference, November 2003.)

[0004] An FFT-based linear equalization algorithm was proposed for
SISO/SIMO systems that provides a good tradeoff between complexity and
performance. The algorithm approximates the non-adaptive LMMSE
algorithm by exploiting the banded Toeplitz structure of the correlation
matrix of the received signal vector. Another attractive alternative is the
Kalman filtering approach proposed in H. Nguyen, K. Zhang, and B.
Raghothaman, "Equalization of CDMA downlink channel via Kalman
filtering." in *Proceedings of Asilomar Conference*, November 2003, where
an interesting state-space model is established to facilitate the application
of the recursive Kalman filter. This approach outperforms the LMMSE
approach at higher complexity.

[0005] For a MIMO system, it has also been shown that both the
conventional LMMSE algorithm and the Kalman filter algorithm can be
extended to the MIMO situation. In these MIMO solutions, both the ICI
and CCI are suppressed in the linear equalization phase of the receiver.
However, these solutions require complicated signal processing whose
computational complexities may be beyond practical limits of the current
hardware implementation. On the other hand, attempts to combine the
non-linear decision feedback interference cancellation with the LMMSE
equalization are also found in the literature, for example, L. Mailander
and J. G. Proakis, "Linear Aided Decision Feedback Equalization of

CDMA Downlink" in Proceedings of Asilomar Conference, November

2003. However, these algorithms perform the decision feedback directly at

the received signal, and thus require the impractical assumption that all the

active Walsh codes are known at the mobile receiver in order to

reconstruct the transmitted chip sequences.

SUMMARY OF THE PREFERRED EMBODIMENTS

[0006] In accordance with an embodiment of the present invention, a

receiver having a first and a second antenna for receiving at least two

spread spectrum symbols from a transmitter having at least first and second

transmit antennas, comprises a first data path for generating a first

estimated symbol $\hat{a}_1(f)$ from the first antenna, a second data path for

generating an estimated symbol sum $\hat{a}_s(f)$ from the first and second

antennas, and an interference cancellation module having inputs coupled to

the first and second data paths, the interference cancellation module for

canceling co-channel interference (CCI) between the estimated symbol sum

and the first estimated symbol to generate a second estimated symbol.

[0007] In accordance with an alternative embodiment of the present

invention, a wireless receiver having at least two receive antennas for

receiving a CDMA transmission from a transmitter having at least two

transmit antennas, comprises a channel estimator having an input coupled

to the at least two receive antennas, a first output, and a second output, a

first chip equalizer having a first input connected to the at least two

receive antennas and a second input of the channel estimator for

suppressing inter-chip interference (ICI) and co-channel interference (CCI)

from at least one antenna other than a first one of the at least two antennas

and for generating an estimated chip sequence from the first antenna, the

first chip equalizer having an output coupled to a first processing module

for descrambling and despreading the output of the first chip equalizer and

generating a first estimated symbol $\hat{a}_1(f)$, a second chip equalizer having a

first input coupled to the at least two receive antennas and a second input

comprising the second output of the channel estimator for generating an

estimated chip sequence sum from the at least two receive antennas and a

residual CCI, the second chip equalizer having an output coupled to a

second processing module for descrambling and despreading the output of

the second chip equalizer and generating an estimated symbol sum $\hat{a}_s(f)$, an

interference cancellation module, having the first estimated symbol $\hat{a}_1(f)$,

the estimated symbol sum $\hat{a}_s(f)$ and an output of the second equalizer as

inputs, for canceling CCI and generating at least one estimated symbol,

and a decoder for decoding the at least one estimated symbols.

[0008] In accordance with an alternative embodiment of the present

invention, a method of receiving a CDMA transmission in a wireless

receiver having at least two receive antennas, the transmission comprising

at least two symbols from a transmitter having at least first and second

transmit antennas, comprises the steps of generating a first estimated

symbol $\hat{a}_1(f)$ from the first antenna, generating an estimated symbol sum

$\hat{a}_s(f)$ from the first and second antennas, and determining a second

estimated symbol by canceling interference between the estimated symbol sum and the first estimated symbol.

[0009] In accordance with an alternative embodiment of the present invention, a wireless receiver having a first and a second receive antennas for receiving a CDMA transmission comprising at least two symbols from a transmitter having at least first and second transmit antennas in which not all spreading codes are known, comprises means for receiving an input data on a first data path for generating a first estimated symbol $\hat{a}_1(f)$ from the first antenna, means for receiving the input data on a second data path for generating an estimated symbol sum $\hat{a}_s(f)$ from the first and second antennas, means for utilizing the first estimated symbol $\hat{a}_1(f)$ and the estimated symbol sum $\hat{a}_s(f)$ as a plurality of inputs to an interference cancellation module, for canceling CCI and generating at least one estimated symbol, and means for decoding the at least one estimated symbol.

[00010]   In accordance with an alternative embodiment of the present invention, a program of machine-readable instructions, tangibly embodied on an information bearing medium and executable by a digital data processor, to perform actions directed toward transmission and receiving methods for multiple antenna technology, the actions comprises receiving as a first input a first estimated symbol $\hat{a}_1(f)$ derived from a first antenna, receiving as a second input an estimated symbol sum $\hat{a}_s(f)$ derived from the first antenna and a second antenna, and calculating a second estimated symbol by canceling

interference between the estimated symbol sum and the first estimated symbol.

[00011]   In accordance with an alternative embodiment of the present invention, a method for receiving a CDMA transmission in a wireless receiver having at least two receive antennas, the transmission comprising at least two symbols from a transmitter having at least first and second transmit antennas, comprises the steps of step for generating a first estimated symbol $\hat{a}_1(f)$ from the first antenna, step for generating an estimated symbol sum $\hat{a}_s(f)$ from the first and second antennas, and step for determining a second estimated symbol by canceling interference between the estimated symbol sum and the first estimated symbol.

BRIEF DESCRIPTION OF THE DRAWINGS

[00012]   These and other features, aspects, and advantages of embodiments of the present invention will become apparent with reference to the following description in conjunction with the accompanying drawings. It is to be understood, however, that the drawings are designed solely for the purposes of illustration and not as a definition of the limits of the invention.

[00013]   Figure 1 is illustrative of an embodiment of a MIMO-CDMA system of the present invention.

[00014]   Figure 2 is a diagram of the transmit signaling at antenna m of an embodiment of the present invention.

[00015]   Figure 3 is a high level block diagram of a typical prior art MIMO-LMMSE receiver.

[00016]   Figure 4 is a block diagram showing an embodiment of a LMMSE-SIC receiver according to the present invention.

[00017]   Figure 5 is a graphical illustration of a mixed traffic 1X EV-DV system.

[00018]   Figure 6 is an illustration of the conditional mean estimator.

[00019]   Figure 7 is a graph showing simulation results of an embodiment of the MIMO LMMSE-SIC algorithm of the present invention.

[00020]   Figure 8 is a graph showing simulation results for the MIMO LMMSE-SIC algorithm with 2 iterations.

DETAILED DESCRIPTION

[00021]   A system according to embodiments of this invention provides a multiple transmit antenna, multiple receive antenna (MIMO) receiver design for communication downlinks such as those used in CDMA technology. Two algorithms referred to as the MIMO LMMSE-FFT and MIMO LMMSE-SIC (Successive Interference Cancellation) algorithms, are described in detail. In embodiments of the invention, the interference cancellation step is achieved without the impractical assumption of the knowledge of all the active Walsh codes in the system.

[00022]   Embodiments of this invention provide a multiple transmit antenna, multiple receive antenna (MIMO) receiver design for communication downlinks such as those used in CDMA technology.

[00023] Multiple transmit, multiple receive antenna MIMO systems offer

potential for realizing high spectral efficiency in a wireless

communications system. Information theoretic studies establish that in an

independent flat-fading channel environment, the capacity of such an

MIMO system increases linearly with the number of antennas. A practical

MIMO configuration, such as a Bell Labs Layered Space-Time (BLAST)

system, may be deployed to realize this high spectral efficiency for a

narrow-band TDMA system. A simpler form of BLAST, the vertical

BLAST (V-BLAST) has been proposed to simplify the transceiver BLAST

design. In V-BLAST, independent parallel data sequences are transmitted

and a successive co-channel interference CCI cancellation/data detection

algorithm is used for efficient multi-symbol detection.

[00024] In the downlink of a CDMA like system, most of the research

has focused on the application of advanced signal processing techniques to

offset the performance degradation due to the loss of Walsh code

orthogonality caused by the frequency selective multipath channel.

Channel equalization is a promising means of improving the receiver

performance in a frequency selective CDMA downlink. Current research

encompasses two approaches to linear equalization, namely non-adaptive

linear equalization (e.g., T. P. Krauss, W. J. Hillery, and M. D. Zoltowski,

"MMSE equalization for forward link in 3G CDMA: Symbol-level versus

chip-level," in Proc. of 10th IEEE workshop on Statistical Signal and

Array Processing, pp. 18–22, 2000), and adaptive linear equalization (e.g.,

M.J. Heikkila, P. Komulainen, and J. Lilleberg, "Interference suppression in CDMA downlink through adaptive channel equalization," in Proc. of VTC 99-Fall, (Amsterdam, The Netherlands), pp. 978–982, 1999, L. Mailaender, "Low-complexity implementation of CDMA downlink equalization," in Proc. of 2001 IEE 3G Mobile Communication Technologies, pp. 396–400, 2001.)

[00025]   Non-adaptive linear equalizers usually assume "piece-wise" stationarity of the channel, and design the equalizer according to some optimization criteria such as Linear Minimum Mean Square Error (LMMSE) or zero-forcing, which in general leads to solving a system of linear equations by matrix inversion. This can be computationally expensive, especially when the coherence time of the channel is short and the equalizers have to be updated frequently.

[00026]   On the other hand, adaptive algorithms solve the similar LMMSE or zero-forcing optimization problems by means of stochastic gradient algorithms and avoid direct matrix inversion. Although computationally more manageable, the adaptive algorithms are less robust since their convergence behavior and performance depend on the choices of parameters such as step size.

[00027]   To overcome these difficulties, in J. Zhang, T. Bhatt, and G. Mandyam, "Efficient linear equalization for high data rate downlink CDMA signaling," in Proceedings of Asilomar Conference, November 2003, an FFT-based linear equalization algorithm was proposed that

provides a good trade off between complexity and performance. The algorithm approximates the non-adaptive LMMSE algorithm by exploiting the banded Toeplitz structure of the correlation matrix of the received signal vector. Another attractive alternative is a Kalman filtering approach wherein an interesting two state-space model is established to facilitate the application of the recursive Kalman filter. This approach outperforms the LMMSE approach at a slightly higher complexity.

[00028]   Applying the MIMO configuration to the CDMA downlink presents a significant challenge to the receiver design, as the receiver has to combat both the ICI and the CCI in order to achieve reliable communication. It has been shown in H. Nguyen, J. Zhang, and B. Raghothaman, "Equalization of CDMA downlink channel via Kalman filtering," (Proceedings of Asilomar Conference, November 2003) that both the conventional LMMSE algorithm and the Kalman filter algorithm can be extended to the MIMO system.

[00029]   In embodiments of the invention, a MIMO LMMSE-SIC algorithm is provided that uses successive interference cancellation to improve the performance of a conventional linear MIMO LMMSE equalizer. One non-limiting advantage of the use of the MIMO LMMSE-SIC algorithm in accordance with embodiments of this invention is that interference cancellation is achieved without the impractical assumption of the knowledge of all the active Walsh codes in the systems, unlike many other interference cancellation based algorithms found in the literature.

The MIMO LMMSE-SIC algorithm detailed herein also incorporates the so-called conditional mean estimator to provide soft decisions in the decision feedback process. Simulation results presented herein suggest that the soft decisions alleviate the error propagation problem in the SIC associated with hard decision feedbacks.

MIMO Signal Model for CDMA Downlink

[00030] Consider an M transmit antenna, N receive antenna MIMO CDMA system as illustrated in Figure 1. Serial to parallel split transmit multiplexing in unit 110 is assumed in order to make the receiver solutions general enough for all possible MIMO transmit multiplexing methods. The modulated symbol stream on line 105 is split at the transmitter into M sub-streams before being transmitted across the M transmit antennas.

[00031] Blocks 115-1 - 115-M represent the process of spreading the data according to a spreading code, such as a Walsh code, and scrambling the data with a known long PN code and allocating the data to the several transmitter antennas 117-1 to 117-M. The signals pass through a channel or channels to receiving antennas 121-1 to 121-N where they are received as signals y_1 to y_N and then detected in a detector/decoder 125 and output on line 127.

[00032] As shown in Figure 2, the signal model at the m_{th} transmit antenna is given as follows, assuming K active Walsh codes in the system:

$$d_m(i) = c(i) \sum_{k=1}^{K} \sum_{m} \alpha_k a_{k,m}(j) s_k(i - jG) \qquad (1)$$

where i, j, m and k are chip, symbol, transmit antenna and spreading code indices respectively. The base station scrambling code is denoted by c(i). Meanwhile, α_k stands for the power assigned to spreading code k (same for all antennas), $a_{k,m}$ is the information symbol sequence for spreading code k at antenna m and s_k is the kth spreading code. Note that in this model it is implicitly assumed that the same set of Walsh codes are used across all the transmit antennas. Signals for the mth antenna having k spreading codes arrive on lines 205-1 to 205-k on the left of Figure 2 and are spread with the k codes in units 210-1 to 210-K, are summed in unit 215 and are processed and scrambled in unit 220, then transmitted along the mth antenna 225. The transmitted signals propagate through the MIMO multipath fading channel denoted by $H_0, ...H_L$, where each matrix is of dimension $N\Delta \times M$, where denotes the number of samples per chip.

[00033]   The N signals are received on N antennas 121-1 to 121-N as signals $y_1, ..._{yN}$, producing y1 $\cdots$ yN RF signals that are detected and decoded in unit 125 and passed out on line 127 for further processing.

[00034]   The signal model at the receive antennas is thus given by the following equation, after stacking up the received samples across all the receive antennas 121-1 to 121-N for the ith chip interval.

$$y_i = \sum_{l=0}^{L} \mathcal{H}_l d_{i-l} + n_i. \tag{2}$$

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Note that $y_i = [y_{i,1}^T, ..., y_{i,N}^T]^T$ is of length $N\Delta$, and each small vector $y_{i,n}$ includes all the temporal samples within the ith chip interval. Meanwhile, L is the channel memory length,

$d_{i-l} = [d_1(i-l), ..., d_M(i-l)]^T$ is the transmitted chip vector at time i-1 and n_i s the $N\Delta \times 1$ dimensional white Gaussian noise vector with

$n_i \sim \mathcal{N}(0, \sigma^2 I)$. Note that σ^2 denotes noise variance and I is the identity matrix. Furthermore, in order to facilitate the discussion on the LMMSE receiver, a block of 2F + 1 received vectors are stacked up:

$$y_{i+F:i-F} = H d_{i+F:i-F-L} + n_{i+F:i-F} \qquad (3)$$

where 2F + 1 is the length of the LMMSE equalizing filter and

$$y_{i+F:i-F} = [y_{i+F}^T, ..., y_{i-F}^T]^T, \quad ((2F+1)N\Delta \times 1)$$

$$n_{i+F:i-F} = [n_{i+F}^T, ..., n_{i-F}^T]^T, \quad ((2F+1)N\Delta \times 1)$$

$$d_{i+F:i-F-L} = [d_{i+F}^T, ..., d_{i-F-L}^T]^T, \quad ((2F+L+1)M \times 1)$$

$$H = \begin{bmatrix} \mathcal{H}_0 & \cdots & \mathcal{H}_L & & \\ & \ddots & & \ddots & \\ & & \mathcal{H}_0 & \cdots & \mathcal{H}_L \end{bmatrix}, \quad ((2F+1)N\Delta \times (2F+L+1)M)$$

where the dimensions of the matrices are given next to them. Note that to keep the notation more intuitive, the subscripts are kept at a block level.

For instance, $y_{i+F:i-F}$ is the vector that contains blocks $y_{i+F}, ..., y_{i-F}$ where each block is a vector of size $N\Delta \times 1$.

MIMO LMMSE Chip-Level Equalization

[00035] The block diagram of the MIMO receiver with chip-level equalizer algorithm is shown in Figure 3. It is a straightforward extension of the LMMSE chip equalization initially designed for a SISO system.

[00036] An input signal received at the various antennas 121 enters on line 305 and passes to channel estimator 310 and chip equalizer 315. The output of the estimator 310 is input to the equalizer so that it may equalize the channels and resolve the various received signals y_1, $\dot{y}_N$. After the chip-level equalizer, the orthogonality of the Walsh code is partially re-installed and all the desired symbols from each transmit antenna are detected with a simple code correlator 320 which correlates to the desired spreading code. The descrambling process is also included in the code correlator block 320, typically prior to de-spreading. Unit 330 performs deinterleaving and decoding.

[00037] Defining an error vector of $z = d_i - W^H y_{i+F:i-F}$ and an error covariance matrix $R_{zz} = E[zz^H]$, the MIMO LMMSE chip-level equalizer W is the solution of the following problem:

$$W^{opt} = \arg\min_{W} \text{Trace}(R_{zz}) = \arg\min_{W} E\|d_i - W^H y_{i+F:i-F}\|^2, \tag{4}$$

whose optimal solution is given by:

$$W^{opt} = \sigma_d^2 R^{-1} H_{i:i}. \tag{5}$$

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Where $R = E[y_{i+F:i-F}y_{i+F:i-F}^H]$ is the correlation matrix of the

received signal, and σ_d^2 is the transmitted chip power. Meanwhile,

although H is fixed for a given channel realization and is not a function of

symbol index i, here there is used the notation $H_{i+F:i+1}$, $H_{i:i}$ and $H_{i-1:i-F-L}$ to

represent the sub-matrices of the overall channel matrix H that are

associated with $d_{i+F:i+1}$, d_i and $d_{i-1:i-F-L}$ in the expansion of the matrix-

vector product:

$$Hd_{i+F:i-F} \triangleq \begin{bmatrix} H_{i+F:i+1} & H_{i:i} & H_{i-1:i-F-L} \end{bmatrix} \begin{bmatrix} d_{i+F:i+1} \\ d_i \\ d_{i-1:i-F-L} \end{bmatrix}$$
$$= H_{i+F:i+1}d_{i+F:i+1} + H_{i:i}d_i + H_{i-1:i-F-L}d_{i-1:i-F-L}. \qquad (6)$$

Constrained Optimization Based Non-linear LMMSE-SIC

[00038]   Both MIMO LMMSE and MIMO LMMSE-FFT belong to the

category of linear equalization methods. In this section, there is introduced

the non-linear decision feedback to the MIMO LMMSE equalizer to

improve the overall receiver performance. The resulting LMMSE-SIC

(LMMSE- Successive Interference Cancellation) receiver is illustrated in

Fig. 4 for a MIMO system having two transmit and N receive antennas.

[00039]   There are two symbol-detection paths 410, 450 in Fig. 4 where

the first/second path aims at detecting symbols transmitted on the

respective first/second transmit antenna. The first path 410 is similar in

certain respects to that of a conventional MIMO LMMSE receiver (eg.,

Fig. 3), where the chip-equalizer 420, having input 421, is connected to antennas 301, 303 and has an output 423 connected to unit 430. Using frequency domain representation for ease of exposition, $\omega_1(f)$ is designed such that both the ICI and the spatial CCI from the other transmit antenna is suppressed in the estimated chip sequence $\hat{d}_1(f)$. This beneficially facilitates the de-spreading and symbol detection in unit 430 to generate a first estimated symbol $\hat{a}_1(f)$ by restoring the orthogonality between the Walsh codes. Note that only the symbols carried on the desired users' Walsh codes are detected. As shown in Figure 4, these detected symbols from the first path are directly fed-back along line 432 and used in the symbol detection of the second path.

[00040]   In the preferred embodiment of the receiver, the channel estimator 440 has an input 441 connected to the antennas 301, 303, a first output 443 connected to equalizer 420 and a second output 442 connected to equalizer 455.

[00041]   The second path 450 of the LMMSE-SIC receiver deviates from that of a conventional MIMO LMMSE receiver at least for the reason that the chip equalizer 455 [$\omega_2(f)$], having input 454 and output 456, does not attempt to directly generate the chip sequence from the second transmit antenna. Instead, it generates a weighted sum of the chip sequences $\hat{d}_s(f)$ on output 456 from both transmit antennas, while suppressing all

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the ICIs. Mathematically,

$$\hat{d}_s(f) = d_2(f) + b_{2,1}d_1(f) + n_2(f). \tag{7}$$

[00042] This filter 455 provides complete temporal ICI suppression while keeping a "controlled" residue CCI whose strength is denoted by a design parameter $b_{2,1}$.Furthermore, this design parameter is jointly optimized with the filter coefficients to achieve the best performance. The detailed derivation of this joint optimization is provided in the next section. Note that $b_{2,1}$ is a post filtering parameter that is obtained by optimizing as in equation (10) below.

[00043]   Since the same set of Walsh codes are used in both transmit antennas, it follows that a simple descrambling/despreading process gives us the symbol sum estimate $\hat{a}_s(f)$ output from despreader 460 (for a particular Walsh code) that also includes a controlled residue CCI:

$$\hat{a}_s(f) = a_2(f) + b_{2,1}a_1(f) + n_2'(f), \tag{8}$$

where the noise variance $\sigma_{n_2'}^2 = 1/G\sigma_{n_2}^2$ and G is the spreading length. Those skilled in the art will appreciate that, assuming correct symbol estimations $\hat{a}_1(f)$ from the first path, the symbol estimate for the second transmit antenna is given by:

$$\hat{a}_2(f) = \hat{a}_s(f) - b_{2,1}\hat{a}_1(f). \tag{9}$$

[00044]   Unit 470 accepts as input the estimated sum from unit 460 and the chip

equalizer function $\omega_2(f)$ (which helps to generate the feedback coefficient $b_{2,1}$ in

box 470), performs the calculation and sends the values of $\hat{\alpha}_1(f)$ and $\hat{\alpha}_2(f)$ to

decoder 330. With the help of a standard soft demodulator (represented

schematically by box 470), the soft channel-coded bits can be easily extracted

from the symbol estimates $\hat{a}_1(f), \hat{a}_2(f)$ before being sent to the decoder along

line 475 to complete the receiver processing. As shown below, significant

performance gain can be achieved with a LMMSE-SIC receiver, compared to a

conventional LMMSE receiver.

[00045]   Furthermore, the LMMSE-SIC algorithm achieves the performance

gains without having to make the impractical assumption of a priori knowledge at

the receiver of all of the active Walsh codes. This important advantage makes the

presently preferred LMMSE-SIC especially attractive for a fixed voice-data

system such as 1X EV-DV or HSDPA and the like, where the desired user usually

accounts for only, typically, ten to fifty percent of the overall transmit power in a

cell. Such a mixed voice-data 1X EV-DV system is illustrated in Figure 5, where

the data user of interest and another data user each consumes about 20 percent of

the transmit power, while the remainder of the transmit power is assigned to voice

users in the cell, in addition to some housekeeping overheads such as pilot,

synchronization, etc. Also note that in this example system, data users have a

spreading length of 32 while the voice users have a spreading length of 64.

Joint Equalizer/Feedback Weights Optimization

[00046]   As noted in the two transmit MIMO example above, the filter

coefficients and the feedback weight b_{12} require joint optimization. In this

section, a joint optimization problem for a general M transmitter MIMO

system is formulated and it is shown that the problem can be converted

into a series of M minimum variance distortionless responses (MVDR)

problems that are easily solved with a Lagrange Multiplier algorithm.

[00047]   Returning to the matrix-vector notation used in equations (3 -6),

the joint equalizer/feedback weights optimization for a M transmit, N

receive MIMO system can be formulated as the following LMMSE

problem with a lower-triangular structural constraint on the feedback

weights:

$$
\mathbf{W}^{opt}, \mathbf{B}^{opt} = \arg\min_{\mathbf{W,B}} \mathrm{Trace}(\mathbf{R}_{zz}) \;=\; \arg\min_{\mathbf{W,B}} E\|\mathbf{B}^{H}\mathbf{d}_i - \mathbf{W}^{H}\mathbf{y}_{i+F:i-F}\|^{2},
$$

$$
\text{s.t.} \quad \mathbf{B} = \begin{bmatrix} 1 & & 0 \\ \vdots & \ddots & \\ b_{M,1} & \ldots & 1 \end{bmatrix}. \tag{10}
$$

Note that in this case the error vector is defined as

$$
\mathbf{z} = \mathbf{B}^{H}\mathbf{d}_i - \mathbf{W}^{H}\mathbf{y}_{i+F:i-F}.
$$

[00048]   The values of the coefficients in the set of chip equalizers (W)

and the feedback weights (B) used to multiply an estimated kth symbol

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(d_i) fed into the calculation of the ith estimated symbol are found by

minimizing the trace of the error correlation matrix, subject to the

constraint on the structure of the matrix B.

[00049]   A direct application of Lagrange multipliers to equation (10)

proves to be difficult. To exploit the structure of the problem, the

structural constraint in (10) is reorganized into two constraints that are

more compact:

$$b_{m,m} = 1, \quad m = 1, \cdots, M \text{ and}$$
$$\mathcal{Z}b = 0 \tag{11}$$

where $b_{m,m} \triangleq (B)_{m,m}$ and

$$b \triangleq vec(B^H) = [b_1^T, ...b_M^T]^T \tag{12}$$

in which b_m is the m_{th} column of B. Note that vec(B) denoted a column-

wise stack-up of the matrix(B). Furthermore, Z is defined as

$$\mathcal{Z} \triangleq \begin{bmatrix} \mathcal{Z}_1 & & \\ & \ddots & \\ & & \mathcal{Z}_M \end{bmatrix} \tag{13}$$

where each sub-matrix Z_m is an M × M diagonal matrix denoting the index

of the null elements in b_i, that is, $(Z_m)_{k,k} = 1$ if the kth element in b_m is

constrained to be zero, and $(Z_m)_{k,k} = 0$ if it is not. Similarly, there is

defined another diagonal matrix $N = \text{diag } N_1 \ldots N_M$ that denotes the index

to the non-zero elements in b. It is evident that $Z_m + N_m = I$ by definition,

where I is the identity matrix. With these definitions, the MMSE problem
is set up to provide optimal W and B with a generalized feedback
structure:

$$
\begin{aligned}
\mathbf{W}^{opt}, \mathbf{B}^{opt} &= \arg\min_{\mathbf{W},\mathbf{B}} E\|\mathbf{B}^H \mathbf{d}_i - \mathbf{W}^H y_{i+F:i-F}\|^2, \quad \text{s.t.}\\
b_{m,m} &= 1, \quad m = 1,\cdots,M \text{ and}\\
\mathcal{Z}\mathbf{b} &= 0.
\end{aligned}
\tag{14}
$$

[00050]   To proceed, note that the expectation in equation (14), referred
to as the cost function J can be written as:

$$
\begin{aligned}
J &= E\|\mathbf{W}^H y_{i+F:i-F} - \mathbf{B}^H \mathbf{d}_i\|^2\\
&= \sum_{m=1}^{M} \|\mathbf{w}_m^H \mathbf{H}_{i:i} - \mathbf{b}_m^H\|^2 + \sum_{m=1}^{M} \mathbf{w}_m^H \mathbf{V}\mathbf{w}_m
\end{aligned}
\tag{15}
$$

where $\mathbf{V} \triangleq \sigma_d^2 \bar{\mathbf{H}}_{i:i} \bar{\mathbf{H}}_{i:i}^H + \sigma^2 \mathbf{I}$, $\mathbf{w}_m$ is the mth column of W. Note that

$\bar{\mathbf{H}}_{i:i}$ is a submatrix of H that is obtained by excluding $\mathbf{H}_{i:i}$ in the matrix H.

Meanwhile, in the derivation, yi+F: i-F has been substituted from equation

(3). The identity $\mathbf{I} = \mathbf{Z}_m + \mathbf{N}_m$ is utilized to break up $\mathbf{H}_{i:i}$ and b_m^H as:

$$
\begin{aligned}
\mathbf{H}_{i:i} &= \mathbf{H}_{i:i}\mathcal{N}_m + \mathbf{H}_{i:i}\mathcal{Z}_m,\\
\mathbf{b}_m^H &= \mathbf{b}_m^H \mathcal{N}_m + \mathbf{b}_m^H \mathcal{Z}_m.
\end{aligned}
\tag{16}
$$

Substituting (16) into (15) yields:

$$
J = \sum_{m=1}^{M} \|\mathbf{w}_m^H \mathbf{H}_{i:i}\mathcal{N}_m - \mathbf{b}_m^H \mathcal{N}_m\|^2 + \sum_{m=1}^{M} \mathbf{w}_m^H \mathbf{V}_m \mathbf{w}_m
\tag{17}
$$

where $V_m \triangleq H_{i:i} Z_m H_{i:i}^H + V$ has been defined. Since $b_m^H \mathcal{N}_m$ contains the

controllable non-zero elements in b_m^H, the first term in J is minimized by setting

$$b_m^H \mathcal{N}_m = w_m^H H_{i:i} \mathcal{N}_m. \tag{18}$$

[00051] Meanwhile, it is observed that the second sum in J is optimized by

individually minimizing

$$w_m^{opt} = \arg\min_{w_m} w_m^H V_m w_m, \quad \text{s.t. } w_m^H h_m = 1; \quad \text{for } m = 1, ..., M \tag{19}$$

Note that h_m is the mth column of the matrix $H_{i:i}$, and the constraint

$w_m^H h_m = 1$ follows from (18) and the original constraint $b_{m,m} = 1$. The

complex optimization problem (10) is thus decoupled into a series of

simple MVDR optimization problems (19) to get the filter coefficients w_m.

The non-zero elements in the feedback weight vectors b_m are then

obtained from (18). A straightforward application of the Lagrange

Multiplier method to (19) leads to the final solution to the joint

optimization problem:

$$\begin{aligned} w_m^{opt} &= \frac{V_m^{-1} h_m}{h_m^H V_m^{-1} h_m} \\ (b_m^{opt})^H \mathcal{N}_m &= (w_m^H)^{opt} H_{i:i} \mathcal{N}_m, \end{aligned} \tag{20}$$

and the elements of the resulting error correlation matrix is given by

$$(R_{zz})_{m,n} = (w_m^{opt})^H (V + H_{i:i} Z_m Z_n H_{i:i}^H) w_n^{opt}. \tag{21}$$

Soft Decision Feedback

[00052]   In the two transmit, two receive example illustrated in Figure 4,

the estimated symbol from the first path $\hat{a}_1(f)$ is used in (9) to obtain the

symbol estimates for the second path. If these symbol estimates $\hat{a}_1(f)$ from

first path are generated by hard-decisions, then the performance

degradation for the second path can be severe due to error propagations.

[00053]   The error propagation problem can be alleviated by using soft

decision feedback. There are many types of soft estimators in the literature.

The so-called conditional mean based estimator is both analytically

pleasing and easy to implement. This conditional mean estimator (also

known as MMSE estimator) is also used in similar decision feedback

algorithms. Consider the general signal model for an arbitrary symbol at

the output of the de-spreader:

$$r_{k,m}(j) = ca_{k,m}(j) + n, \tag{22}$$

where c is a multiplicative factor and k, m, j are Walsh code, transmit

antenna and symbol indices, respectively. Also note n is assumed to be

Gaussian: $n \sim \mathcal{N}(0, \sigma_n^2)$ and both c and σ_n^2 are functions of the spreading

gain G, transmit power α_k, noise level σ^2, channel H and the filter Wopt.

Once c and σ_n^2 are obtained, the conditional mean estimate of $a_{k,m(j)}$ is

given by

$$\hat{a}_{k,m}(j) = E[a_{k,m}(j)|r_{k,m}(j)], \tag{23}$$

which has a closed-form solution if one invokes the Gaussian assumption. For

example, when the symbols of interest are QPSK modulated with unit energy, the

solution reduces to a simple hyperbolic tangent function:

$$\hat{a}_{k,m}(j) = \frac{1}{\sqrt{2}} tanh\left(\frac{Re(\sqrt{2}c^{*}r_{k,m}(j))}{\sigma_{n}^{2}}\right) + j\frac{1}{\sqrt{2}} tanh\left(\frac{Im(\sqrt{2}c^{*}r_{k,m}(j))}{\sigma_{n}^{2}}\right). \qquad (24)$$

[00054]   The reason that the conditional mean estimator works well can

be intuitively explained by observing Figure 6, where the real part of the

symbol estimate $Re(\hat{a}_{k,m}(j))$ is plotted as a function of the real part of the

initial noisy soft estimate $Re(r_{k,m}(j))$. The hyperbolic function

essentially acts as a clipper, which provides near-hard decision output

when the initial estimate is reliable, i.e, when $|Re(r_{k,m}(j))|$ is large. On

the contrary, if the initial estimate is less reliable, the output is reduced to

alleviate the possible error propagation impact.

Optimization of Detection Order and Further Iterations

[00055]   In the LMMSE-SIC algorithm discussed above, the detection

order is assumed to be (1, 2, ...,M). However, the performance of the

successive detection can be improved by optimizing the detection order.

The detection order for a similar V-BLAST problem can be chosen so that

the worst SNR among M data streams is maximized. In particular, let

$\omega = (\omega_{1}, \cdots, \omega_{M})$ denote an arbitrary ordering and let Ω be the set of all

possible orderings. The cardinality of the set is $|\Omega| = M!$ and the

following problem is solved to obtain the optimal ordering β:

$$\beta = \arg\max_{\omega \in \Omega} \ \min_{m=1}^{M} \gamma_m(\omega), \tag{25}$$

where $\gamma_m(\omega)$ denotes the effective SNR associated with the detection of the mth

transmitted data sequence for a given detection order ω. The effective SNR can

be equivalently defined either at chip-level or at symbol level. The chip-level SNR

for the mth data sequence is defined as a function of the error covariance matrix

R_{zz}:

$$\gamma_m(\omega) = \frac{\sigma_d^2}{(R_{zz}(\omega))_{m,m}}, \tag{26}$$

which reduces to the following simple form considering the results in equations

(20) and (21):

$$\gamma_m(\omega) = \sigma_d^2 h_m^H V_m^{-1}(\omega) h_m. \tag{27}$$

[00056]   Once the effective SNR is defined, equation (25) can be solved

by a localized optimization procedure. In a localized optimization

procedure, at each stage of the detection process, the sequence that has the

best SNR is chosen for detection.

[00057]   The performance of the LMMSE-SIC algorithm can be improved

if more complexity is allowed and further iterations are introduced after all

the data sequences are detected using the successive algorithm described

above. Without loss of generality, it is assumed that in the first iteration

the detection order is $w = (1, 2,,,..., M)$ and the filter matrix W and

feedback matrix B is obtained by solving the optimization equation (10). In the second iteration, however, the design of the matrix pencil W,B should be changed to reflect the fact that at the beginning of the second iteration, there has already been obtained an initial estimate of all the transmitted data sequences. To this end, the constraint in equation (10) is changed and the optimization equation rewritten as:

$$\mathbf{W}^{opt}, \mathbf{B}^{opt} = \arg\min_{\mathbf{W},\mathbf{B}} E\|\mathbf{B}^H \mathbf{d}_i - \mathbf{W}^H y_{i+F:i-F}\|^2;$$

$$\text{s.t.} \quad \mathbf{B} = \begin{bmatrix} 1 & \ldots & b_{1,M} \\ \vdots & \ddots & \vdots \\ b_{M,1} & \ldots & 1 \end{bmatrix}. \tag{28}$$

[00058]   It is observed that in equation (28), none of the elements in the feedback matrix B is constrained to be zero. Therefore, it allows the maximum amount of decision feedback in the spatial dimension and thus improves the effective SNR for each data sequence. Note that since equation (28) can also be transformed into the form shown in equation (14) with the compact constraint representation, the solution in (20-21) applies to (28) as well. However, it is easy to see that in this case,

$$\mathcal{Z}_1 = \cdots = \mathcal{Z}_M = 0, \tag{29}$$

and as a result

$$\mathbf{V}_1 = \cdots = \mathbf{V}_M = \mathbf{V}. \tag{30}$$

[00059]   Consequently, the solution to (28) assumes a simpler form:

$$
\begin{aligned}
\mathbf{w}_m^{opt} &= \frac{\mathbf{V}^{-1}\mathbf{h}_m}{\mathbf{h}_m^H\mathbf{V}^{-1}\mathbf{h}_m} \\
(\mathbf{b}_m^{opt})^H &= (\mathbf{w}_m^H)^{opt}\mathbf{H}_{i:i}.
\end{aligned}
\tag{31}
$$

[00060]   One further denotes γ_m^{II} as the effective SNR for the mth data

sequence in the second iteration. It is easy to show that:

$$
\gamma_m^{II} = \sigma_d^2 \mathbf{h}_m^H \mathbf{V}^{-1}\mathbf{h}_m.
\tag{32}
$$

[00061]   Note that unlike $\mathbf{V}_m$, $\mathbf{V}$ is independent of the detection order.

[00062]   Consequently, both $\mathbf{w}_m$, $\mathbf{b}_m$, and γ_m^{II} are independent of the

detection order, meaning that all these quantities can be computed at the

same time, unlike the local optimization process required in the first

iteration. The detection order in the second iteration is thus obtained by

simply sorting all the γ_m^{II} in descending order. Furthermore, denoting

γ_m^{I} as the SNR for the mth data sequence in the first iteration, one can

easily show that:

$$
\gamma_m^{I} \leq \gamma_m^{II},
\tag{33}
$$

for any $1 \leq m \leq M$. Note that this inequality holds regardless of the

detection order used in the first and second iterations. This inequality also

signifies the benefit of having a second iteration in the LMMSE-SIC

algorithm, provided that the added complexity is within the confines of a

practical implementation.

FFT-based Complexity Reduction for a 2 Tx LMMSE-SIC Algorithm

[00063]   The complexity of the MIMO LMMSE algorithm can be greatly
reduced with a FFT-based approach that exploits the block Toeplitz
structure of the received signal correlation matrix R, and approximates $\mathbf{R}^{-1}$
with $\mathbf{S}^{-1}$ where S is the associated block circulant matrix. Because the
matrices to be inverted in the LMMSE-SIC algorithm, namely $V_{m,m} = 1, \cdots$
$,M$, are not block Toeplitz, the direct extension of the FFT-based approach
to the LMMSE-SIC algorithm is not possible for a general M transmit, N
receive MIMO system. For the special case where the number of transmit
antennas $M = 2$, the matrices V_1 and V_2 can be written as a rank-one or
rank-two update of a block Toeplitz matrix, which allows a low-
complexity inversion that is similar to the MIMO LMMSE-FFT approach.
If one focuses on the first iteration of LMMSE-SIC and assumes that the

detection order is $\omega = (1, 2)$.

[00064]   One can show that in this case:

$$V_1 \;=\; V + \sigma_d^2 h_2 h_2^H$$

$$V_2 \;=\; V. \tag{34}$$

and they relate to the correlation matrix R by

$$V_1 = R - \sigma_d^2 h_1 h_1^H$$

$$V_2 = R - \sigma_d^2 H_{t:i} H_{t:i}^H = R - \sigma_d^2 h_1 h_1^H - \sigma_d^2 h_2 h_2^H. \tag{35}$$

[00065]   Substituting (35) into (20) and using the identity that

$$\frac{A^{-1} c}{c^H A^{-1} c} = \frac{(A - \sigma_d^2 cc^H)^{-1} c}{c^h (A - \sigma_d^2 cc^H)^{-1} c}, \tag{36}$$

[00066]   One can write the optimal filters as a function of correlation

matrix R:

$$w_1 = \frac{V_1^{-1} h_1}{h_1^H V_1^{-1} h_1} = \frac{R^{-1} h_1}{h_1^H R^{-1} h_1},$$

$$w_2 = \frac{V_2^{-1} h_2}{h_2^H V_2^{-1} h_2} = \frac{(R - \sigma_d^2 h_1 h_1^H)^{-1} h_2}{h_2^H (R - \sigma_d^2 h_1 h_1^H)^{-1} h_2}. \tag{37}$$

[00067]   It is clear in (37) that the FFT-based approach can be used

directly to get w_1 by approximating R with its associated circulant matrix

S: $R^{-1} \approx S^{-1}$. On the other hand, the application of the FFT approach to

w_2 requires a little more work. Observing that $R - \sigma_d^2 h_1 h_1^H$ rank-one

update of the block Toeplitz matrix R, a matrix inversion Lemma is

invoked and there is obtained:

$$(R - \sigma_d^2 h_1 h_1^H)^{-1} = R^{-1} + \frac{\sigma_d^2 R^{-1} h_1 h_1^H R^{-1}}{1 - \sigma_d^2 h_1^H R^{-1} h_1} \approx S^{-1} + \frac{\sigma_d^2 S^{-1} h_1 h_1^H S^{-1}}{1 - \sigma_d^2 h_1^H S^{-1} h_1}. \tag{38}$$

[00068]   Note that the approximation $R^{-1} \approx S^{-1}$ has been used in (38).

From equations (37 -38), it has been shown that for a two transmit MIMO

system using LMMSE-SIC algorithm, the filters w_1 and w_2 can be obtained

using FFT-based approach to compute the inversion of the associated

circulant matrix S, instead of the direct inversion of the correlation R. As a

result, the complexity of the MIMO LMMSE-SIC algorithm for a two

transmit system can be close to that of the MIMO LMMSE-FFT algorithm,

provided that no further iterations are used in the LMMSE-SIC.

Simulation Results

[00069]   The presently preferred MIMO LMMSE-SIC algorithm was

evaluated in a realistic simulation chain, and the simulation parameters are

tabulated in Table I.

Parameter Name	Parameter Value
System	CDMA 1X/EVDV
Spreading Length	32
Channel Profile	Vehicular A
Mobile Speed	50 km/h
Filter Length	32
Number of Tx/Rx Antennas	2/2
Modulation Format	QPSK
Information Data Rate	312 kbps
Turbo Code Rate	0.6771
Geometry	6
Number of Walsh Codes Assigned to the user	3
Total number of Active Walsh Codes in the system	25

Table 1: Simulation parameters.

[00070]   The benefit of the non-linear decision feedback is demonstrated

in Figure 7, where the MIMO LMMSE-SIC algorithm is compared with the

conventional LMMSE algorithm. Note that the impact of detection order

optimization is included in these results. There are three LMMSE-SIC

curves shown in this figure, representing the hard-decision feedback method, the conditional mean estimator based soft decision feedback method and the ideal feedback case. It is observed that even with the hard decision feedback, the MIMO LMMSE-SIC algorithm outperforms the conventional LMMSE by about 1 dB at high SNR. Another 0.5-1dB can be gained if the hard decisions are replaced with the soft decisions generated by the conditional mean estimator. It is significant that these gains are achieved with the assumption that the mobile receiver has only the knowledge of 3 out of 25 active Walsh codes.

[00071]   In Figure 8, the impact of further iterations in the LMMSE-SIC algorithm is investigated using the same example. Note that for both the one iteration and two iterations cases, the conditional mean estimator based soft decision feedback is used. In this case, adding a second iteration actually degrades the performance at high SNR. This detrimental effect, also known as "ping-pong" effect in the iterative detectors, is mainly attributed to the accumulation of decision errors in the iterative process. Although described in the context of particular embodiments, it will be apparent to those skilled in the art that a number of modifications and various changes to these teachings may occur. Thus, while the invention has been particularly shown and described with respect to one or more preferred embodiments thereof, it will be understood by those skilled in the art that certain modifications or changes, in form and shape, may be

made therein without departing from the scope and spirit of the invention as set forth above.